CHAPTER 5

IMC BASED CONTROLLER FOR LOAD DISTURBANCE REJECTION

5.1 INTRODUCTION TO INTERNAL MODEL BASED CONTROLLER

The drawback of conventional PID controllers is that it generally does not work well for non-linear systems and particularly complex and vague systems that have no precise mathematical models. To overcome the drawback, various types of modified conventional PID controllers were developed lately. Some well-known and simple methods are hysteresis control, stator-frame proportional integral (PI) control and synchronous-frame PI control. Hysteresis control allows only a fixed switching frequency and Stator-frame PI control leads to control error at steady state but Synchronous-frame PI control has been accepted as being superior that yields zero control error at steady state.

IMC was originally developed for Chemical Engineering applications by Morari and Zafiriou (1989) and is considered as a robust control method. In IMC, Synchronous-frame PI controllers are obtained and the controller parameters are expressed directly in certain machine parameters and the desired closed-loop bandwidth. This simplifies the design procedure and trial-and-error steps are avoided. Structure of IMC is shown in Figure 5.1. The control system includes two blocks, namely controller and model. Model is the perfect representation of the process. The control system inputs are set
point and process output and the output is the manipulated variable. The parallel path with this model is to subtract the effect of manipulated variable from the process output. Feedback signal is equal to the influence of disturbances, and it is not affected by the action of the manipulated variables. Hence, the system is effectively open loop and it is not affected by the stability problems associated with feedback control systems. It cancels the influence of disturbances because the feedback signal is equal to this influence and modifies the controller set point accordingly.

![Figure 5.1 Structure of IMC based controller](image-url)

**Figure 5.1 Structure of IMC based controller**

Since the difference between the plant and the reference model is used as feedback, dynamic and steady errors caused by immeasurable disturbances and uncertainty of control process are reduced by IMC controller. Also, by defining a suitable reference model, regulation of controller parameters online is possible with robustness.

In the IMC structure of Figure 5.1, $G(s)$ is the plant, $\hat{G}(s)$ is the reference model, $C_{\text{imc}}(s)$ is the internal model controller, and $D(s)$ is the disturbance. The feedback signal obtained for IMC is given below.
where, $I$ is the identity matrix. For a perfect reference model where $\hat{G}(s) = G(s)$, the feedback signal is equal to the disturbance $D(s)$. When there are differences between the plant and the model, $E(s)$ provides the information of disturbance and also model–plant mismatch, and hence the robustness is obtained by compensating for the derivation appropriately. IMC control scheme is mathematically analogous to conventional feedback controller as given below.

$$F(s) = \frac{C_{\text{imc}}(s)}{I - C_{\text{imc}}(s)\hat{G}(s)}$$  \hspace{1cm} (5.2)$$

When the reference model is accurate, the stability of the overall system depends on the stability of both controller and plant. The system control error can be minimized by making $C_{\text{imc}}(s)\hat{G}(s) = I$ under the assumption that $\hat{G}(s) = G(s)$ and that $G(s)$ is stable. When the IMC controller satisfies, $C_{\text{imc}}(0)\hat{G}(0) = I$, a unit step change in $R(s)$ will yield zero steady-state error for any asymptotically constant disturbance $D(s)$.

### 5.2 Parameter Estimation of IMC Based Current Controller

Consider the complex space vector equations for the IM in synchronous coordinates as presented by Lennart Harnefors and Hans-Peter Nee (1998),

$$\frac{d\psi_s(t)}{dt} = -R_s i_s(t) - j\omega_s \psi_s(t) + \nu(t),$$  \hspace{1cm} (5.3)$$
Here, \( \bar{\psi}_S, \bar{\psi}_S \) and \( \bar{\psi}_R, \bar{\psi}_R \) are the stator and rotor current and flux space vectors respectively, while \( \bar{v} \) is the impressed stator voltage space vector. \( R_s, R_r, \) and \( L_s, L_r \) are the stator and rotor resistances and self-inductances respectively. \( L_m \) is the magnetizing inductance, while \( \omega_1 \) and \( \omega_2 = \omega_1 - \omega_r \) are the stator and slip frequencies, respectively. After eliminating \( \bar{i}_r \) and \( \bar{\psi}_s \) among the above equations, it is obtained that,

\[
L \sigma \frac{d\bar{i}_s(t)}{dx} + \left( R_s + \left( \frac{L_m}{L_r} \right)^2 R_r \right) \bar{i}_s(t) + j\omega_1 L \sigma \bar{i}_s(t) = \bar{v}(t) + \frac{L_m}{L_r} \left( \frac{R_r}{L_r} - j\omega_1 \right) \bar{\psi}_r(t)
\]

where, \( R_{IM} = \left[ R_s + \left( \frac{L_m}{L_r} \right)^2 R_r \right] \), \( L_{\alpha} = L_s - \frac{L_r^2}{L_r} \)

But rotor flux orientation implies that \( \bar{\psi}_r \) is real and constant. Even though perfect field orientation never can be accomplished in practice, \( \bar{\psi}_r \) will atleast vary slowly and can be considered as a quasi-constant “disturbance” which is compensated for by integral control action. Hence, including the flux term in a modified stator voltage vector \( \bar{v}' \), the stator current vector is written as,
\[ \bar{I}_r(t) = \frac{1}{(s + j \omega_1)L_\sigma + R_{IM}} \bar{V}(s) \] (5.8)

In the Equation (5.8), \((s + j \omega_1)L_\sigma + R_{IM} = \bar{G}(s)\). The complex-valued transfer function \(\bar{G}\) corresponds to the inverse transfer function matrix

\[
G^{-1}(s) = \begin{bmatrix}
  sL_\sigma + R_{IM} & -\omega_1 L_\sigma \\
  \omega_1 L_\sigma & sL_\sigma + R_{IM}
\end{bmatrix} \] (5.9)

Thus, for the IMC based current controller shown in Figure 5.2 for the rotor-flux-oriented IM control is given by,

\[
F(s) = \frac{1}{s} \begin{bmatrix}
  sL_\sigma + R_{IM} & -\omega_1 L_\sigma \\
  \omega_1 L_\sigma & sL_\sigma + R_{IM}
\end{bmatrix}
\] (5.10)

from which \(F_{pf}(s)\) for decoupling and diagonal IMC or standard PI control is obtained by putting \(\omega_1 = 0\).

**Figure 5.2** Block diagram of IMC based current controller
5.3 PARAMETER ESTIMATION OF IMC BASED SPEED CONTROLLER

For the design of the speed controller as shown in Figure 5.3, the Internal Model Control method has been used. The mechanical dynamics are expressed in electrical radians as presented by Sünter and Altun (2004),

\[
\frac{d\omega_r}{dx} = \frac{1}{J} \left[ \frac{P}{2} (T_m - T_i) - B\omega_r \right] \tag{5.11}
\]

The developed torque can be expressed as

\[
T_m = \frac{2}{3} P I_m \frac{L_m}{L_R} \lambda_r = \frac{2}{3} P P_L \frac{L_m}{L_R} \left( q s \lambda_d - i ds \lambda r \right) \tag{5.12}
\]

Substituting the torque equation in the Equation (5.11), it gives the following Equation (5.13),

\[
\omega_r = \frac{1}{J s + B} \left( \frac{P}{2} k_i \left( i_q - \frac{T_i}{k_i} \right) \right) \tag{5.13}
\]

where, \( s \) is the Laplace operator and \( k_i = 3\psi_n P/2 \) is the torque constant in Nm/A.

![Figure 5.3 Block diagram of IMC based speed controller](image)

Figure 5.3 Block diagram of IMC based speed controller
Since the electrical dynamics are much faster than the mechanical dynamics, with an accurate current controller, it is possible to set \( i_q = i_{q*} \) and take \( i_{q*} \) as an input signal. Assuming the load torque effect as a disturbance, a speed controller can be designed using the method described in the design of current controller. The result obtained is given below.

\[
k_{ps} = \frac{\alpha_s J}{k_i P} k_{is} = \frac{\alpha_s B}{k_i P}
\]

(5.14)

where \( \alpha_s \) is the bandwidth of the speed controller. The friction coefficient \( B \) varies with speed, temperature and weight on the shaft, and is hard to determine. Therefore, this is not a good property, and it causes a drawback in the design procedure, since the integration term \( k_{is} \) is a function of the friction coefficient. If \( B \) was assumed to be zero in the design procedure, the result is simply a P-controller and the integration term is needed to remove the remaining errors and to dampen out oscillations. The proportional term \( k_{ps} \) is not a function of \( B \), so an active damping can be introduced and \( i_{qref} \) is written as,

\[
i_{qref} = i_{qref}^* - \frac{b_o}{k_i} \omega_r
\]

(5.15)

Here, \( \frac{b_o}{k_i} \omega_r \) is the active damping term.

The transfer function from \( \omega_r \) to \( i_{qref}^* \) can be written as

\[
G_s = \frac{\omega_r}{i_{qref}^*} = \frac{k_i \frac{p}{2}}{J s + B + b_o \frac{p}{2}}
\]

(5.16)
To simplify $G(s)$, the bandwidth is set to $\alpha_s = \left( \frac{b_\omega J P}{2} \right) + (B/J)$.

Solving for $b_a$ yields

$$b_a = \frac{\alpha_s J - B}{P} \frac{2}{P}$$

(5.17)

Assuming as $J >> B$, the transfer function $G_s$ is simplified to

$$G_s = \frac{k \left( \frac{P}{2} \right)}{s + \alpha_s}$$

(5.18)

Therefore, the PI parameters can be determined from

$$G_{pi} = \frac{\alpha_s}{s} G^{-1}(s)$$

(5.19)

$$k_{ps} = \frac{\alpha_s J}{k_i P} \quad \text{and} \quad k_{is} = \frac{\alpha_s^2 J}{k_i P^2}$$

(5.20)

In this work, the rise time of the speed was chosen to be fifty times slower than that for the current controllers, since the speed loop is added as an outer loop to the current loop.

### 5.4 IMPLEMENTATION OF VECTOR CONTROL TECHNIQUE USING IMC

The block diagram of vector control technique implemented using the proposed IMC based controllers is shown in Figure 5.4. The induction motor is supplied through three phase to three phase direct AC to AC matrix
converter. Inner current control and outer speed control loops with IMC based controllers are used for the implementation of vector control technique. Apart from the three popular modulation algorithms which are discussed in the previous chapter, the harmonics present in input current, output voltage and current as well as waveform distortion are very less with space vector modulation algorithm and hence it is used to control the matrix converter switching. Actual three phase currents of the motor $i_a$, $i_b$, and $i_c$ are measured using current sensors and converted into two phase stationary reference frame variables $i_\alpha$ and $i_\beta$ using Clarke’s transformation, presented by Toliyat and Campbell (2004) as in Figure 5.5.

Figure 5.4 Block diagram of IMC based MC fed IM drive system
Figure 5.5 Clarke’s Transformation

From the Figure 5.5, the following equations are developed.

\[
\begin{bmatrix}
i_a \\
i_b \\
i_c
\end{bmatrix} = \begin{bmatrix}
cos\theta & sin\theta \\
\cos(\theta-120) & \sin(\theta-120) \\
\cos(\theta+120) & \sin(\theta+120)
\end{bmatrix} \begin{bmatrix}
i_\alpha \\
i_\beta \\
i_0
\end{bmatrix}
\] (5.21)

The third component of the second matrix in the above Equation (5.21) is zero sequence components and it will not make any difference in adding. From the corresponding inverse transformation of the above Equation (5.21), the two phase stationary reference frame variables are calculated from the measured three phase currents of the motor.

\[
\begin{bmatrix}
i_\alpha \\
i_\beta \\
i_0
\end{bmatrix} = \begin{bmatrix}
cos\theta & \cos(\theta-120) & \cos(\theta+120) \\
sin\theta & \sin(\theta-120) & \sin(\theta+120) \\
0.5 & 0.5 & 0.5
\end{bmatrix} \begin{bmatrix}
i_a \\
i_b \\
i_c
\end{bmatrix}
\] (5.22)

Figure 5.6 Clarke’s Transformation with \( \theta=0 \)
For convenience, if $\theta$ is made to zero and neglecting zero sequence components as shown in Figure 5.6, the transformation equations can be simplified and written as,

$$i_a = \frac{2}{3} i_a - \frac{1}{3} i_b - \frac{1}{3} i_c = i_a$$  \hspace{1cm} (5.23)$$

and

$$i_\beta = -\frac{1}{\sqrt{3}} i_b + \frac{1}{\sqrt{3}} i_c$$  \hspace{1cm} (5.24)$$

The two phase stationary reference frame variables $i_a$ and $i_\beta$ are then be converted into two phase rotating reference frame variables $i_d$ and $i_q$ with reference to Figure 5.7.

![Figure 5.7 Transformation between reference frames](image)

**Figure 5.7 Transformation between reference frames**

The transformation equations relating two phase stationary and rotating reference frame variables are given as,

$$i_d = i_\beta \sin \theta_e + i_a \cos \theta_e$$  \hspace{1cm} (5.25)$$

and

$$i_q = i_\beta \cos \theta_e - i_a \sin \theta_e$$  \hspace{1cm} (5.26)$$

In figure.5.7, $\bar{I}$ is defined as,

$$i_a = i_m \cos(\theta_e + \phi)$$  \hspace{1cm} (5.27)$$
\[ i_b = i_m \cos(\theta_e - 120 + \phi) \quad (5.28) \]
\[ i_b = i_m \cos(\theta_e + 120 + \phi) \quad (5.29) \]

Using the above equations, the two phase rotating reference frame variables can also be written as

\[ i_d = i_m \cos \phi \quad (5.30) \]

and

\[ i_q = -i_m \sin \phi. \quad (5.31) \]

The above transformations are well applied to voltage and flux linkage also. The Clarke and Park transformation are used to obtain the actual value of two phase rotating reference frame value of the motor. Desired reference values are generated through IMC based speed and current control loops and also by using inverse Park’s and inverse Clarke’s transformation, reference voltage vector is generated. From this reference voltage vector, switching pulses are generated using space vector modulation algorithm. Finally, the switching pulses are applied to the bidirectional switches of matrix converter.

5.5 SIMULATION RESULTS

Mathematical models of matrix converter, control algorithm and dynamic d-q model of induction motor are implemented using Matlab/Simulink software. The vector control technique is applied to induction motor using IMC based controllers for inner current control and outer speed control loops. In Figures 5.8 to 5.9, the four quadrant operation of matrix converter fed vector controlled induction motor drive with PI and IMC controllers are described. From 0 to 0.5 s, the drive is operated in first quadrant as motor, at rated speed and torque. At t=0.5 s, a step change in speed reduction in the same direction is initiated, the converter transfers the
operation from first quadrant to second quadrant which involves braking at the maximum torque from initial speed to the desired speed is reached, the converter then transfers the operation back to motoring and the drive settles at a point where the motor torque decreases and then becomes equal to the load torque.

**Figure 5.8** Four quadrant operation of vector controlled IM with PI controller

**Figure 5.9** Four quadrant operation of vector controlled IM with IMC
At $t=1.5$ s, speed reference is changed to speed reversal. The converter transfers the operation to second quadrant and the motor braked to zero speed at maximum torque. At zero speed, the converter transfers the operation to reverse motoring. Then the motor is accelerated at maximum torque in the reverse direction in third quadrant until the desired speed is reached. Finally, the drive settles at a point where the motor torque decreases and then becomes equal to load torque. At $t=2.5$ s, the speed reference is changed again to positive direction of rotation. The drive passes through fourth quadrant with maximum torque and settles in first quadrant with desired speed and steady torque. In all these set point reference changes, the settling time with PI controller is approximately 0.5 s and it is reduced to 0.1 s with IMC controller. Hence, the fast transient response of the drive is achieved with IMC based controllers. In Figure 5.10, the speed response of the drive with a change in load torque applied to the motor with IMC controller is shown. The results show that the drive settles faster without any speed overshoot with IMC controller.

![Figure 5.10 Speed and torque response of the drive with load torque disturbance applied at $t=0.2$ s, with IMC](image)
5.6 SUMMARY

Internal model control based speed and current controllers are modeled and implemented with vector controlled induction motor drive fed by matrix converter. The four quadrant operation of the drive with IMC controller is compared with PI controller. The settling time in all the step change in speed with IMC controller is found to be less as compared to PI controller. A step change in load is applied to the drive and the performance is noticed. A fast set point tracking and load disturbance rejection without any overshoot in speed is achieved with IMC control method as compared to PI controller. Hence, the perfect set point tracking and disturbance rejection with better performance of the drive using IMC control method is verified in this work. This work is presented in International Journal of Electrical Power and Energy Systems. A compensation strategy of matrix converter under distorted input voltage conditions is presented in the next chapter.